Design Concepts for Microwave GaAs FET Active Filters

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Abstract —**GaAs MMIC technology has already been employed to great advantage in the design of various active circuits, for example, matching networks, multipliers, and circulators. The development of active *filters* in this technology seems to us to represent a logical step forward in the evolution of GaAs MMIC circuit applications. This paper gives a comprehensive history of microwave active filter development to date, and then presents a design methodology for the realization of precision broad-band filters in spite of problems such as the nonideal behavior of microwave GaAs FET's and the low Q and related parasitic effects of MMIC inductors and capacitors. Finally, the feasibility of our approach is illustrated by the computer-simulated design of a cascadable, second-degree microwave bandpass filter.**

I. Introduction

ANALOG FILTERS in the microwave frequency regime have enjoyed wide application in communications, radar, and signal processing systems. Traditionally, such filters have been implemented as passive networks of waveguide, transmission-line, or discrete lumped elements. More recently, microwave filters have also been realized in surface acoustic wave (SAW) and magnetostatic surface wave (MSW) technology. The design methodology for these networks is relatively mature; excellent summaries of the various techniques are given in, for example, [1]–[5]. The past few decades have also seen the invention of the transistor and the development of the integrated circuit, along with the miniaturization of wide varieties of electrical networks. In particular, at frequencies up to about 1 MHz, analog filters have undergone a great reduction in size with the advent of active-filter technology. (See, for example, [6].) These active filters are realizable as *hybrid* integrated circuits of resistors, capacitors, and op amps or as *monolithic* integrated circuits containing switches, capacitors, and op amps. Inductors are not used because it is virtually impossible to simultaneously miniaturize these components, minimize their dissipation, and maintain usable inductance values. However, in the microwave frequency region, filters are still constructed in distributed or discrete lumped-parameter form, and consequently remain quite bulky, especially in comparison to the microwave monolithic integrated circuits (MMIC's) now being used for amplifiers and other nonfiltering circuits. While high-power networks will continue to require the more traditional waveguide and transmission-line style implementations, signal processing and related applications would clearly benefit from the availability of MMIC filter realizations.

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The use of MMIC technology to realize microwave filters would clearly be advantageous because of the potentially great performance-to-size ratio that could be achieved, and the compatibility of such networks with other well-established MMIC signal processing circuits. However, MMIC filters of even moderate precision and selectivity could not be built using passive elements alone because of the undesirable dissipation and parasitic effects of these components. (See, for example, [7, p. 93], where a five-section Chebyshev bandpass filter modeled using ideal capacitors and inductors with Q's of 25 at 10 GHz results in a passband insertion loss of 5 to 6 dB.) Hence if MMIC filters are to be realized at all, they would have to be built using both lumped components *and* GaAs FET's, where the gain of the active devices would compensate somehow for the lossiness of the passive elements. However, GaAs FET's themselves are nonideal, and their parameters depend in a complicated way upon frequency, biasing, and signal level. Hence an active microwave filter would have to contain sufficient degrees of freedom for it to achieve a prescribed filter response, despite the imperfections of both the passive *and* active components. Other problems that would have to be dealt with include minimizing the noise figure, guaranteeing stability, reducing effects of parasitics and process sensitivity, minimizing sensitivity to variations in the GaAs FET parameters, and maximizing dynamic range while limiting nonlinearities. However, inasmuch as these difficulties have largely been overcome in the design of MMIC *amplifiers*, we feel confident that MMIC active *filters* could also be successfully designed. Indeed, the next section discusses some very encouraging preliminary results along these lines already achieved by us and other researchers.

It should be pointed out that the great success of active filters at frequencies below 1 MHz is due to the mature integrated-circuit technology that has produced inexpensive, high-performance, and virtually imperfection-free *operational amplifiers*. Such devices are not yet available at microwave frequencies. Indeed, we are faced with different constraints upon the usable circuit elements in the microwave regime, and this prevents us from directly employ-

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ing low-frequency active-filter design methods to produce active MMIC filters.

II. Previous Work on Microwave Active Filters

The compelling advantages of active MMIC's are reflected in the considerable amount of recent work in this area, in particular for active *circulators* [8], [9], active *multipliers* [10], active *gyrators* [11], and active *inductors* [12]. Regarding microwave active filters, various efforts have been made over the past 20 years to develop these networks, although most of this research has not emphasized circuit realization in MMIC technology. We now present a summary of this previous work, along with a review of a closely related area: active *matching network* design.

A. Simulated High-Q Inductors and Capacitors

One of the earliest references to microwave active filters was made in 1969, when Adams and Ho [13] described a technique for using bipolar transistors directly as high-Q inductors in the UHF region (300–1000 MHz). Their idea was to take advantage of certain transit-time properties of nonideal microwave transistors so as to realize an inductive effect. By so doing, they were able to construct resonators consisting of a simulated inductance shunted by a lumped capacitance which were then used in bandpass filters. This approach allowed the realization of miniaturized, high-selectivity filters for UHF. To achieve comparable selectivity in passive UHF filters would require the use of high-Q lumped or distributed inductors, which invariably causes the network to be much larger than the active realization of Adams and Ho. Some further discussion of their method is given in [14] and [15].

A considerable amount of work on microwave active filters has been performed in the Soviet Union. The design philosophy adopted there is similar to that of Adams and Ho, although the implementation is somewhat more elaborate. Namely, Soviet researchers have adopted the approach of simulating dissipation-free reactive elements by employing transistors in conjunction with lossy capacitors and inductors. The resulting circuits are then used as the resonant elements in conventional filter structures. In [16] this approach is implemented by using negative immittance converters to introduce a negative resistive component in series (parallel) with the inductors (capacitors) so as to cancel the passive-element dissipation. (This technique has also been used in varactor-tunable filters [17].) In [18] this same approach is implemented using a Q-enhancement technique. With this method, signal energy is fed into a passive, low-Q resonant circuit, and the resulting output is coupled into a transistor which amplifies and feeds back power to the tuned circuit. The fed-back power is arranged to be in phase with the input, and the power gain is set equal to the dissipation of the tuned circuit. This allows the effective Q of the circuit to become arbitrarily large.

In the approaches described above, the broad-band behavior of the active devices used with the nonideal reactive elements is *not* generally taken into account. This has limited the applicability of these methods to narrow-band cases only.

B. A Microwave Active All-Pass Network

In our own work [19], we established the feasibility of *broad-band* MMIC active filters. Specifically, we designed, in a computer simulation, an all-pass network consisting of GaAs FET's and lumped *RLC* elements and possessing the transfer function

$$S_{21}(s) = \frac{s^2 - (\omega_0/q)s + \omega_0^2}{s^2 + (\omega_0/q)s + \omega_0^2}$$

with $f_0 = \omega_0/2/\pi = 10$ GHz and $q = 3$. First, the network was designed using ideal elements. Next, the frequency-dependent effects of microwave GaAs FET's were introduced, which naturally destroyed the response accuracy of the filter. Then, we demonstrated that we could compensate for the nonideal properties of the GaAs FET's and restore the response accuracy over 8–12 GHz by *optimizing* the lumped element values of the filter. This work was significant because it illustrated that it is possible to obtain a prescribed broad-band gain and phase response for an active microwave filter network despite the imperfections of the active devices. Although such design by optimization had already been widely applied to microwave *amplifiers*, we believe that we were the first to apply the technique to microwave *filters*.

In Section IV we present the computer-simulated design of an active microwave bandpass filter. In that example, we consider not only the imperfections of the active devices, but also the effects of bias circuitry and reactive element losses. We also show that it is possible to cascade filter sections to realize a higher order response.

C. Transversal and Recursive Filters

In [20], Rauscher applies transversal and recursive filter principles to the design of microwave active filters. This ingenious methodology allows the realization of broad-band filters comprising transistors, transmission lines, and lumped *RLC* components. Because distributed elements are used, however, this technique—as presently implemented—yields filters which are necessarily compatible in size with *hybrid* microwave integrated circuits. Nevertheless, it is suggested in [20] that the filters could be adapted for realization as MMIC's by replacing the transmission lines by appropriate configurations of GaAs FET's and passive elements. Such an approach would appear to be quite promising.

D. An Active Gyrator

The approach taken in [12] represents a most useful step in the development of microwave active filters. In this work, an inductance was created by terminating an MMIC active gyrator with a capacitor. A narrow-band (0.86–

0.92 GHz) bandpass filter with a Q approaching 200 at 0.9 GHz was then realized by shunting the simulated inductor with a (second) capacitor. Elaborations of this technique have been used with great success to design audio-frequency active filters, and it is most interesting that the method has been extended—at least in an elementary way—to microwaves.

E. *Active Matching Network Design*

In a microwave amplifier, matching networks are used to manage the transfer of power between the source, the active devices, and the load. Customarily, the elements comprising matching networks are *passive*. However, the benefits of active MMIC technology are substantial, and consequently, researchers have developed *active* matching network realizations. (See, for example, [21]–[24].) Common-gate FET circuits are generally used to implement active input and output matching stages. Actively matched amplifiers are smaller than comparable passively matched realizations, and are more tolerant of component-value variations. Even further circuit miniaturization has been achieved by using active inductances, as reported in [12].

Matching networks and filters are similar, since matching networks provide a prescribed passband response between *complex* impedances, while filters provide prescribed passband and stopband responses between *real* resistances. This, along with the fact that matching networks have already been implemented in active MMIC technology, provides substantial motivation for the development of active microwave filters.

III. A Proposed Circuit Implementation for Microwave Active Filters

There are several possible approaches that can be taken for designing microwave active filters. One might attempt to adapt the circuits and techniques that have proved successful at low frequencies, or to use some novel approaches. In either case, two basic shortcomings of MMIC components must be overcome: the frequency dependence of the GaAs FET's, and the low Q and parasitics of the lumped elements.

Of the many ideas that could be considered, the *cascade* approach attempts to localize and simplify the detrimental effects of the nonideal MMIC components by realizing a complicated filter as a cascade of several simpler, noninteracting networks. This approach is quite straightforward, and has, in fact, been used with great success in the realization of audio-frequency active filters. There are other low-frequency design methods which better emulate the low-sensitivity properties of passive LC filters, and which are consequently superior to the cascade approach, especially for highly selective networks. These techniques currently depend upon the availability of high-quality operational amplifiers and eventually might become applicable to microwave filters. In our judgment, the cascade approach, by virtue of its relative simplicity, represents a reasonable starting point for MMIC filter design.

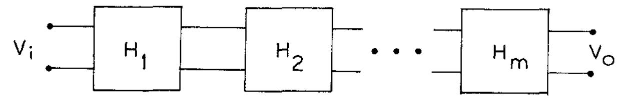

Fig. 1. Cascaded biquadratic filter sections.

A. *Cascade Realization*

The idea is to start with a transfer function $H(s)$ which is to be realized as a microwave active filter:

$$H(s)=\frac{p_0+p_1s+p_2s^2+\cdots+p_ns^n}{e_0+e_1s+e_2s^2+\cdots+e_ns^n}=\frac{V_{\text{out}}}{V_{\text{in}}}.$$

(The p_i and e_i are real constants; s is the complex frequency variable.) This function is then factored as a product of $m=n/2$ biquadratic factors:

$$H(s)=\prod_{i=1}^{m}\frac{a_is^2+b_is+c_i}{g_is^2+h_is+k_i}=\prod_{i=1}^{m}H_i(s).$$

(The a_i, b_i, c_i, g_i, h_i, and k_i are real constants.) Note that we have implicitly assumed above that n is even; naturally, if n is odd there will be $(n-1)/2$ biquadratic factors and one bilinear factor which may, with no loss of generality, be considered as a limiting case of a biquadratic factor. Assume, then, that the filter is realized as a cascade of m elementary two-ports as shown in Fig. 1.

Now, if the transfer function of the ith two-port is $H_i(s)$, and if interconnecting the two-ports as shown does not affect their transfer functions, then clearly

$$H(s)=H_1(s)H_2(s)\cdots H_i(s)\cdots H_m(s).$$

Thus the problem of realizing an arbitrary nth-degree transfer function is reduced to the considerably simpler task of being able to build a general *second*-degree filter section that does not detrimentally interact with other such elementary sections when placed in a cascade configuration. Such noninteraction may be achieved by arranging the input impedance of each section to be very *high* and the output impedance to be very *low*; this is normally the case with audio-frequency active filters. At microwave frequencies, a more practical solution is to design each section so its input (output) impedance is 50 Ω when the corresponding output (input) is terminated in 50 Ω. This technique, which allows the sections to be cascaded without detrimental interaction, was first implemented in the classical passive, *constant-resistance* networks developed over 50 years ago [25].

B. *Circuit Structures*

We have identified a number of transistor networks that can realize biquadratic transfer functions and hence have potential for cascade microwave filter realization, e.g. [26]–[29]. These circuits, shown in Fig. 2(a)–(d) in their original bipolar form, either are derived from negative impedance converter (NIC) circuits, or are based upon active realizations of passive constant-resistance networks. (These networks were, in fact, used to realize low-frequency active filters before op amps supplanted transistors in

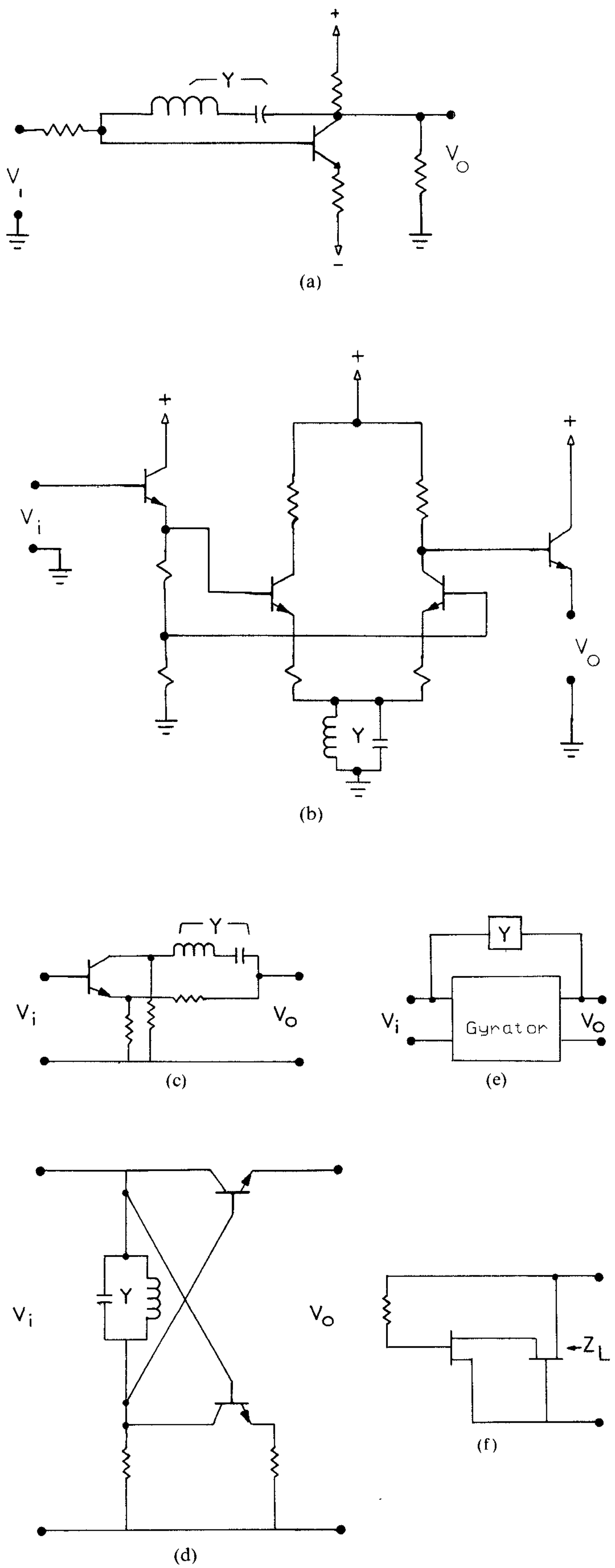

Fig. 2. Transistor networks for active filter applications: (a) Rubin–Even all-pass. (b) Two-transistor Orchard all-pass. (c) One-transistor Orchard all-pass. (d) Larky all-pass. (e) Gyrator-based filter. (f) Active inductor [12].

these applications.) With ideal components, they each can realize the transfer function

$$\frac{V_{out}}{V_{in}} = \frac{Y - G_0}{Y + G_0}$$

where Y is the admittance indicated in each circuit of Fig. 2(a)–(d), and G_0, with the dimensions of conductance, is a function of the other elements of the respective circuits. As is well known from circuit theory, an *all-pass* response results whenever Y is purely reactive (i.e., lossless), and a low-pass, high-pass, bandpass, or band-reject response can result from the appropriate selection of a lossy admittance for Y. For use at microwave frequencies, each circuit would have to be implemented using GaAs FET's and MMIC lumped elements. Another circuit capable of realizing the above transfer function is shown in Fig. 2(e), where a feedback admittance is bridged across a gyrator. In [30] it is shown that this network displays the desirable constant-resistance property discussed above, which allows a cascade connection of such two-ports without detrimental interaction between the individual sections. Since an MMIC gyrator has already been demonstrated in [12], this network is certainly an excellent candidate for microwave filter realization.

C. The Design Task

The fundamental filter design problem is to find a circuit structure with component values in response to a specification of a filter transfer function. If any of the circuits of Fig. 2 are to be used in the cascade approach described above, the first step would be to create a microwave version of the network by replacing any existing bipolar transistors with GaAs FET's. The passive elements also would have to be modeled to include MMIC parasitics. At this point it would also be appropriate to identify those circuits which are better suited to high-Q or low-Q applications. For example, it is known that NIC-based circuits are conditionally stable, while gyrator-based active filters are absolutely stable, so the latter group would undoubtedly be better for more stringent filtering requirements. The remaining steps of the design process may be enumerated as follows.

1) After factoring the overall filter transfer function, design the corresponding biquadratic circuits to provide the desired frequency response. Each section must provide its specified response in spite of the nonideal properties of the active and passive MMIC components.
2) Design the sections—possibly by obtaining constant-resistance behavior—so that the act of cascading the networks does not deleteriously affect the overall frequency response of the filter.
3) Obtain passive components values for the sections that are realizable with current technology.
4) Within the inherent constraints of the cascade approach, achieve a design that is not unduly sensitive to variations in the component values of the filter or the device parameters, and that meets other require-

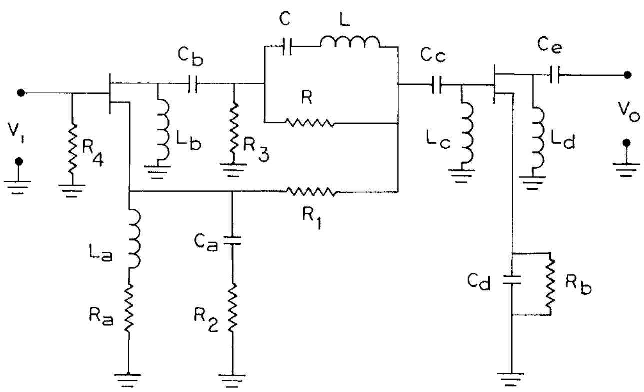

Fig. 3. Microwave bandpass filter section with bias circuitry and finite-Q inductors and capacitors.

ments involving noise figure, dynamic range, stability, and so on.

Although it would be preferable to formulate a deterministic design procedure, it is likely that iterative optimization techniques will be necessary because of the nonideal MMIC components used in the filters. This should not present much difficulty since computer optimization is routinely employed in MMIC amplifier design, which is closely related to the present problem.

In the next section we illustrate, in a computer simulation, the design of a cascadable, second-degree microwave filter section. The main purpose of this exercise is to illustrate how at least *some* of the steps of the proposed design procedure may be implemented. It is also hoped that the simulation will underscore the feasibility of the cascade design approach for MMIC filters, and will provide a stimulus for further research in this area.

IV. Computer-Simulated Design Example

A. The Design Problem

We first wish to design a bandpass filter which, between 50 Ω terminations, realizes the magnitude of the transfer function

$$S_{21}(s) = T(s) = \frac{(\omega_0/q)s}{s^2 + (\omega_0/q)s + \omega_0^2} \tag{1}$$

with $f_0 = \omega_0/2/\pi = 2$ GHz and $q = 1$. We then wish to realize the magnitude of the transfer function

$$S_{21}(s) = T(s) * T(s) \tag{2}$$

by cascading two of the networks realized in the first part of this example. Fig. 3 shows the circuit to be used for this relatively low-Q second-degree bandpass. It is derived from the networks shown in Fig. 2(c) and [19], and it additionally includes bias circuitry $L_a - L_d$; $C_a - C_e$; R_a and R_b. (The bias sources have been set to zero.) Each inductor and capacitor is modeled with a finite Q.

B. Single-Stage Design Method

The approach taken is similar to that used for designing broad-band microwave amplifiers. We start with an initial idealized design which is then perturbed by introducing component imperfections. Computer optimization is then applied to restore the frequency response to the desired shape.

We first consider an idealized version of the circuit of Fig. 3. The following assumptions are made.

1) The bias elements have no effect upon the frequency response. That is, the inductances $L_a - L_d$ and the capacitances $C_a - C_e$ are assumed to be infinite.
2) The reactive elements L and C are lossless.
3) The transistor nearest the input is modeled as an ideal voltage-controlled current source with transconductance g_m.
4) The transistor nearest the output is modeled as an ideal unity-gain buffer.
5) $(1/R_1 + 1/R_3) \gg g_m$; $R_2 = R_3$; $R = R_1$; and $(1/R_4) = 0$.

$S_{21}(s)$ in Fig. 3 is then easily derived as

$$S_{21}(s) = \frac{2V_o}{V_i} = -2\frac{[R/(2L)]s}{s^2 + [R/(2L)]s + 1/LC}.$$

The center frequency ω_0 and q defined in (1) are then given by

$$\omega_0 = 1/\sqrt{LC} \quad \text{and} \quad q = (2/R)\sqrt{L/C}.$$

A center frequency of 2 GHz and a q of 1 may be achieved by selecting $L = 3.166$ nH, $C = 2$ pF, and $R = R_1 = 79.58$ Ω. The values of R_2 and R_3 must be set equal to each other but are otherwise arbitrary; we select a value of 100 Ω.

The ideal design serves merely to select a reasonable set of initial values in the nonideal circuit of Fig. 3. For this network, we let the active devices be characterized by the manufacturer's specified scattering parameters over 0.1–8 GHz for a packaged NE71084 GaAs FET. For the bias circuitry, $L_a - L_d$ were each set to 100 nH; $C_a - C_e$ were each set to 100 pF; and R_a and R_b were set to 12 Ω. These values are typical of what might be required to bias the NE71084's. R_5 was set equal to 50 Ω, and the Q's of all the reactive elements were set and maintained at 25.

The next step was to optimize the values of R, $R_1 - R_4$, L, and C so as to achieve the response specified by (1). The imperfections of the active devices, the lossy reactive elements, and the bias circuitry caused a significant error in the frequency response. We compare the desired response of (1) to the response before optimization in Fig. 4. The indicated frequency range of 0.5–8 GHz, which covers two octaves below and above the center frequency of 2 GHz, was chosen as a realistic bandwidth over which the design goals might be met. In addition, it was decided to match the desired gain shape, rather than the absolute values of the gain across the band, since we felt that the circuit would be able to provide significant gain at the center frequency. Our own program CiAO [31] was used to perform a Fletcher–Powell gradient optimization, and the

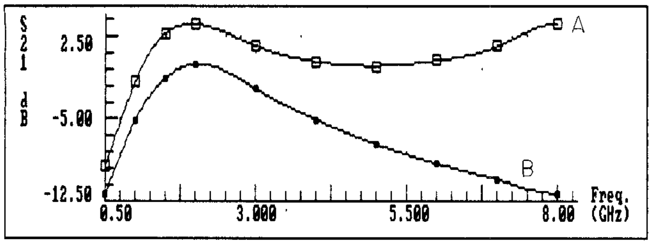

Fig. 4. A: frequency response of circuit of Fig. 3 with initial set of element values B: desired response for circuit of Fig. 3.

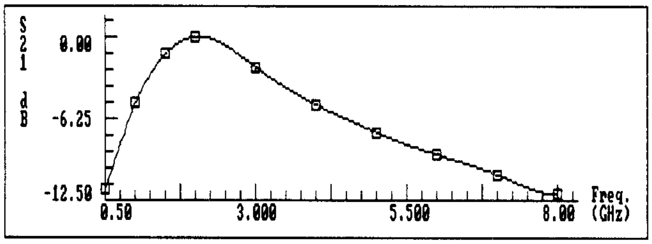

Fig. 5. Response shape after optimization of circuit of Fig. 3. Response error within ± 0.2 dB across the band.

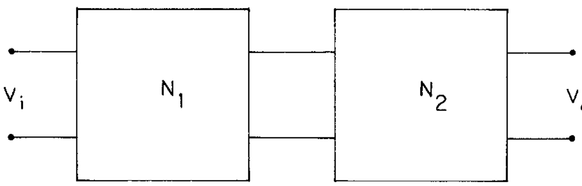

Fig. 6. The cascade of two networks of the type shown in Fig. 3.

results for the element values were as follows:

$$
\begin{aligned}
R &= 309.8\ \Omega \\
R_1 &= 61.97 \\
R_2 &= 15.46 \\
R_3 &= 154.4 \\
R_4 &= 39.28 \\
L &= 8.675\ \text{nH} \\
C &= 0.5881\ \text{pF}.
\end{aligned} \tag{3}
$$

The response error was ± 0.2 dB across the band, with a gain of 9.6 dB at 2 GHz. S_{21} (normalized to 0 dB gain at 2 GHz) for the optimized circuit is shown in Fig. 5.

C. Cascaded Sections

To achieve the transfer function (2) by cascading two identical filters of the type designed above would require that there be no deleterious interaction between the constituent sections. This could be achieved if each second-degree circuit displayed constant-resistance behavior, as mentioned earlier. Although the design was not specifically performed to obtain this, we noted that when two identical sections were cascaded, the response shape came within 2 dB of that given by (2) over 2–8 GHz. This motivated the following strategy for the cascade design. In the first section, denoted by N_1 in Fig. 6, the elements were fixed at the values listed in (3). For the second section, N_2, the element values were optimized to obtain the response shape given by (2). The results for N_2 were as follows:

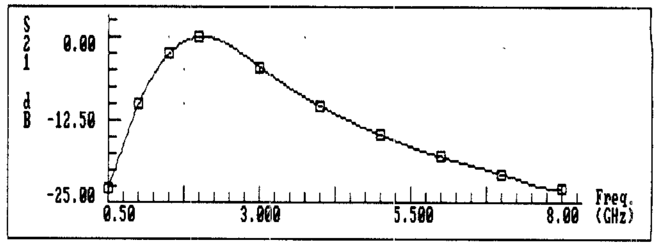

Fig. 7. Response shape after optimization of two cascaded circuits of the type shown in Fig 3. Response error within $+0.16/-1.06$ dB across the band.

$$
\begin{aligned}
R &= 386.1\ \Omega \\
R_1 &= 61.88 \\
R_2 &= 15.19 \\
R_3 &= 162.5 \\
R_4 &= 29.06 \\
L &= 9.270\ \text{nH} \\
C &= 0.5505\ \text{pF}.
\end{aligned} \tag{4}
$$

The response error was $+0.16/-1.06$ dB across the band ($+0.16/-0.66$ dB over 1–8 GHz), with a gain of 16.5 dB at 2 GHz. S_{21} (normalized to 0 dB gain at 2 GHz) for the optimized circuit is shown in Fig. 7.

D. Discussion

It is interesting to compare the response of a simple series LC circuit designed to provide the same response as (1) between 50 Ω terminations. Using $L = 7.958$ nH, $C = 0.7958$ pF, and Q's of 25 for both elements, the passive circuit realized 0.7 dB loss at the center frequency of 2 GHz. While this may not be significant for a single second-order section, the cumulative effect of such element dissipation in higher order filters would clearly be unacceptable. This agrees with the conclusion of [7] already mentioned by us in Section I. We also examined the sensitivity of our second-order circuit to variations in the FET S parameters by calculating the response using the manufacturer's specified parameters for an unpackaged device. The overall response shape changed by only a few tenths of a dB. Our circuit was also observed to be stable across the entire band.

V. Conclusion

The development of MMIC active filters represents an important step forward in the evolution of totally integrated microwave systems. In many applications, such filters would replace the relatively bulky waveguide and transmission-line filters with miniature circuits that provide equivalent performance. Presented herein has been a summary of active microwave filter development to date, along with ideas for further research. In particular, a computer-simulated design example has illustrated the feasibility of the cascade filter-synthesis approach. It would appear that active microwave filters may be no more difficult to design than microwave amplifiers, although further research is needed to examine the effects of noise, nonlinearities, process variations, and the like.

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